A Surface Equivalence-Based Method to Enable Rapid Design and Layout Iterations of Coupled Electromagnetic Components in Integrated Packages

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Abstract

A novel methodology is presented that expedites the electromagnetic analysis in the design cycle of individual layout components in close proximity to other radiating and electromagnetic structures. The proposed method retains all the advantages of a surface based moment method technique, but avoids explicit modeling of the interactions between the object under design and the neighboring ones, without compromising on the accuracy of capturing the electromagnetic coupling between them. As a result the simulation time in individual design cycle is greatly reduced.

I. Introduction

"Faster" and "smaller" are two key foci of today's electronic industry; the operating frequency or switching speeds and the packing density are increasing in modern day packages. As a result, full-wave electromagnetic and proximity effects are playing increasingly more important roles in determining the package performance. Also, low power applications reduce the noise tolerance and therefore demand greater fidelity in modeling the aforementioned effects especially in densely coupled environments. As a result accurate electromagnetic analysis of complex coupled system is of paramount interest in today's packaging technology.

In a package design process, the final layout is achieved through a number of design iterations where the electrical behavior of the layout is numerically modeled at each iteration step. In a densely packed system, such computation becomes very expensive due to the coupling effects from the interacting objects, such as mutually coupled inductors, located in close proximity to the component under design.

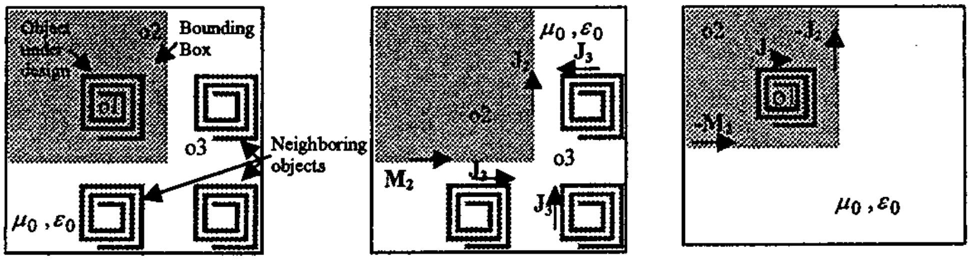
Finite element methods [1] are based on the differential form of Maxwell's equations and lead to a sparse matrix system of equations. This system also has the advantage that different components are represented independently in a decoupled manner, thereby enabling design. FEM based methods do suffer from some disadvantages that include large system sizes due to volumetric discretization, frequency dependent meshing to capture exact skin effects, and the requirement for accurate absorbing boundary conditions (ABCs) to truncate FEM meshes. As an alternative, surface integral equation based techniques such as the Method of Moments (MoM)[2] have become popular in package-level EM simulation due to smaller system sizes, automatic built in radiation conditions (and hence no ABCs), and frequency independent meshing upto foreseeable frequencies of interest. For design purposes, however, the MoM poses a challenge. The Green's function-based underlying formulation results in a highly dense and highly coupled system, where every piece of the discretization interacts with every other piece through a mutual coupling term. Circuit-centric variations of the MoM such as the Partial Element Equivalent Circuit (PEEC) [3] method also suffer from the similar problem, as all the circuit elements representing the cross-coupling between the discretized representation of a design-iterated object and the discretized representation of unaltered objects in close proximity need to be recomputed and re-solved at each design iteration step. The reason for the inherent inefficiency is the inability to exploit the unchanging nature of interacting components that have already been designed or are fixed in design.

The proposed method retains the strengths of surface based MoM formulations while removing the bottleneck of explicitly modeling the coupling between the geometry under design and all the neighboring layout components in every design iteration. The paradigm is based on using the surface equivalence principle [4] on an enclosing mathematical surface, followed by storing the Schur complements [5] of the sub-set of the system matrix that do not change with the design iterations. It is important to note that the proposed technique bypasses the cost of modeling all the mutual coupling explicitly without compromising on the accuracy in accounting for the coupling.

II. Formulation

Isolation

The electromagnetic isolation of the object under design is achieved by introducing a smooth mathematical surface around the object. The surface isolates the problem into an exterior equivalent problem (Fig.1b) and an interior equivalent problem (Fig. 1c), where the object under design belongs only to the interior problem and the neighboring layout components belong to the exterior problem.



(a) (b) (c)

Fig.1 (a) Original problem (b) Exterior equivalent problem (c) Interior Equivalent problem

It is important to notice that the exterior equivalent problem does not change with a change in the design or location of the object o1 within the mathematical surface. However if the shape and size of the object o1 change drastically, it may require a change in the smooth surface o2, to maintain a desired level of accuracy. As a result of decomposing the actual problem into the equivalent problems the mutual couplings between the object under design (o1) and the other neighboring objects (o3) are captured in two separate problems and are never required to be modeled explicitly.

Figure 1 depicts the equivalent surface currents in both exterior and the interior problems for a simple case where all the physical objects (o1 and o3) are modeled as perfect electric conductors. However the methodology presented here is generally valid for any object where the material property is modeled using surface impedance approximations or more rigorous lossy dielectric models [6]. Here the metal objects o1 and o3 support equivalent electric currents $\mathbf{J}_1$ and $\mathbf{J}_3$ respectively, whereas the mathematical bounding box o2 supports both electric current $\mathbf{J}_2$ and magnetic current $\mathbf{M}_2$. On the metal surface, the electric field integral equation [2] is enforced, whereas on the dielectric surface, the continuity of tangential electric and magnetic fields across dielectric boundary is enforced, as in a PMCHWT formulation [4]. Thus combining all the interactions the overall system of linear equations is written in a matrix form as

$$\begin{bmatrix} \mathbf{E}_{\mathbf{J}_1}^{1,int} & -\mathbf{E}_{\mathbf{J}_2}^{1,int} & -\mathbf{E}_{\mathbf{M}_2}^{1,int} & 0 \\ -\mathbf{E}_{\mathbf{J}_1}^{2,int} & \mathbf{E}_{\mathbf{J}_2}^{2,ext}+\mathbf{E}_{\mathbf{J}_2}^{2,int} & \mathbf{E}_{\mathbf{M}_2}^{2,ext}+\mathbf{E}_{\mathbf{M}_2}^{2,int} & \mathbf{E}_{\mathbf{J}_3}^{2,ext} \\ -\mathbf{H}_{\mathbf{J}_1}^{2,int} & \mathbf{H}_{\mathbf{J}_2}^{2,ext}+\mathbf{H}_{\mathbf{J}_2}^{2,int} & \mathbf{H}_{\mathbf{M}_2}^{2,ext}+\mathbf{H}_{\mathbf{M}_2}^{2,int} & \mathbf{H}_{\mathbf{J}_3}^{2,ext} \\ 0 & \mathbf{E}_{\mathbf{J}_2}^{3,ext} & \mathbf{E}_{\mathbf{M}_2}^{3,ext} & \mathbf{E}_{\mathbf{J}_3}^{3,ext} \end{bmatrix} \begin{bmatrix} J_1 \\ J_2 \\ M_2 \\ J_3 \end{bmatrix} = \begin{bmatrix} -\mathbf{E}_1^{inc} \\ -\mathbf{E}_2^{inc} \\ -\mathbf{H}_2^{inc} \\ -\mathbf{E}_3^{inc} \end{bmatrix} \quad (1)$$

Where $\mathbf{X}_{\mathbf{Y}_\beta}^{\alpha,k}$ in general represents the tangential electric or magnetic field $\mathbf{X}$ ($\mathbf{E}$ or $\mathbf{H}$) on the surface α (o1,o2 or o3), due to the source current $\mathbf{Y}$ (electric current $\mathbf{J}$ or magnetic current $\mathbf{M}$) on surface β (o1,o2 or o3), radiating in the k (interior or exterior) region. The unknown vector represents the strengths of the electric and magnetic currents, and the right hand side vector (RHS) consists of the incident electric or magnetic field in the corresponding media, tangential to the surface. Typically for an external radiation-free microelectronic environment there is no direct excitation on the surface o2, but circuit excitations (modeled here as simple gap-source excitations) [7] exist for the appropriate basis functions on the surfaces o1 and o3.

At each step of the design iteration for the object o1, only the blocks associated with the surface of o1, i.e. the blocks in the first row and the first column change. However, since there is no direct interaction between the surfaces of o1 and o3, the corresponding blocks are null. Typically, there is a large number of unknowns associated with o3 that represents the union of all the discertizations of surfaces of objects interacting with o1. Thus the isolation prevents the larger size matrices involving basis functions on o3 from changing through the design iterations.

Matrix Decomposition

The linear system in (1) can be represented in a more convenient and compact way by representing the unknowns on o1, o2 and o3 by $\mathbf{Y}_1$, $\mathbf{Y}_2$ and $\mathbf{Y}_3$, and all the tested fields by $\mathbf{X}^1$, $\mathbf{X}^2$ and $\mathbf{X}^3$, as

$$\begin{bmatrix} \mathbf{X}^1_{\mathbf{Y}_1} & \mathbf{X}^1_{\mathbf{Y}_2} & 0 \\ \mathbf{X}^2_{\mathbf{Y}_1} & \mathbf{X}^2_{\mathbf{Y}_2} & \mathbf{X}^2_{\mathbf{Y}_3} \\ 0 & \mathbf{X}^3_{\mathbf{Y}_2} & \mathbf{X}^3_{\mathbf{Y}_3} \end{bmatrix} \begin{bmatrix} \mathbf{Y}_1 \\ \mathbf{Y}_2 \\ \mathbf{Y}_3 \end{bmatrix} = \begin{bmatrix} \mathbf{R}_1 \\ 0 \\ \mathbf{R}_3 \end{bmatrix} \tag{2}$$

where $\mathbf{R}_1$ and $\mathbf{R}_3$ represents the excitation.

Finally using variable substitution, the unknown on the object under design, i.e. o1, can be written as

$$\mathbf{Y}_1 = \left[\mathbf{X}^1_{\mathbf{Y}_1} - \mathbf{X}^1_{\mathbf{Y}_2}\{\mathbf{X}^2_{\mathbf{Y}_2} - \mathbf{X}^2_{\mathbf{Y}_3}(\mathbf{X}^3_{\mathbf{Y}_3})^{-1}\mathbf{X}^3_{\mathbf{Y}_2}\}^{-1}\mathbf{X}^2_{\mathbf{Y}_1}\right]^{-1}\left[\mathbf{R}_1 + \mathbf{X}^1_{\mathbf{Y}_2}\mathbf{X}^2_{\mathbf{Y}_3}(\mathbf{X}^3_{\mathbf{Y}_3})^{-1}\mathbf{R}_3\right] \tag{3}$$

Clearly, the terms $\{\mathbf{X}^2_{\mathbf{Y}_2} - \mathbf{X}^2_{\mathbf{Y}_3}(\mathbf{X}^3_{\mathbf{Y}_3})^{-1}\mathbf{X}^3_{\mathbf{Y}_2}\}^{-1}$, and $\mathbf{X}^2_{\mathbf{Y}_3}(\mathbf{X}^3_{\mathbf{Y}_3})^{-1}$ do not change during the design iterations of object (o1), and can be pre-computed for all the design steps.

If N_1, N_2, N_3 are the number of unknowns on the surface o1, o2, o3, the cost of solving for $\mathbf{Y}_1$ during the stages of the design iterations where the appropriate blocks are pre-computed is given by C_1, where

$$C_1 = N_1^3 + N_1 N_2 (N_1 + N) + N_1^2 + N_1 N_2 + N_2 N_3 \tag{4a}$$

If the flexibility of changing the excitation on the neighboring geometry is not required the cost can be further reduced to C_1' given by $C_1' = N_1^3 + N_1 N_2 (N_1 + N) + N_1^2 + N_1 N_2$ (4b)

The direct approach without any isolation or decomposition has the cost given by C_2, where

$$C_2 = (N_1 + N_3)^3 + (N_1 + N_3)^2 \tag{5}$$

Also, it is straightforward to identify the unchanging nature of the neighboring object (o3). Therefore using just decomposition without the isolation the solution $\mathbf{Y}_1$ can be computed as

$$\mathbf{Y}_1 = \left[\mathbf{X}^1_{\mathbf{Y}_1} - \mathbf{X}^1_{\mathbf{Y}_3}(\mathbf{X}^3_{\mathbf{Y}_3})^{-1}\mathbf{X}^3_{\mathbf{Y}_1}\right]^{-1}\left[\mathbf{R}_1 + \mathbf{X}^1_{\mathbf{Y}_3}(\mathbf{X}^3_{\mathbf{Y}_3})^{-1}\mathbf{R}_3\right] \tag{6}$$

Where the cost is given by C_3, where $C_3 = N_1^3 + N_1 N_3 (N_1 + N_3) + N_1^2 + N_1 N_3 + N_3^2$ (7)

For accurate analysis of a small component in the layout we need to consider the coupling effects due to all the nearby components, i.e. $N_3 >> \{N_1, N_2\}$. Thus from (4,5 and 7) we conclude for such cases $C_1 << \{C_2, C_3\}$.

III. Numerical Results

The presented method is applied in extracting the scattering (S) parameters of a transmission line structure(Fig. 2a) in the presence of a near-by spiral inductor and a ground plane. The structures are residing in a medium with relative dielectric constant of 4. In the design iteration, where the line width is changed from $3\mu m$ to $5\mu m$, and the gap between the two conductors is reduced from $10\mu m$ to $8\mu m$, the new S-parameters are modeled using the isolation technique presented in the paper. The result is in excellent agreement (Fig. 2b) with that obtained by a fully coupled analysis of the perturbed design including all the neighboring objects. Table 1 demonstrates the ability of the proposed isolation technique in modeling the coupling effects from the nearby objects, for a 4 by 4 spiral inductor array (Fig. 3a). It is important to note that the coupling effects increase as frequency is increased. Also, the S-parameters corresponding to ports 1 and 2s behave asymmetrically due to the uneven nature of coupling from the neighboring inductors, as opposed to the case where the coupling is not modeled where the S-parameters are symmetric.

IV. Conclusion

The proposed methodology develops and utilizes a surface-equivalence based isolation technique to accelerate repeated electromagnetic analysis through the design iterations of a layout component in a densely coupled environment. The accuracy and the relative cost advantage of the proposed method have been demonstrated compared to the existing techniques.

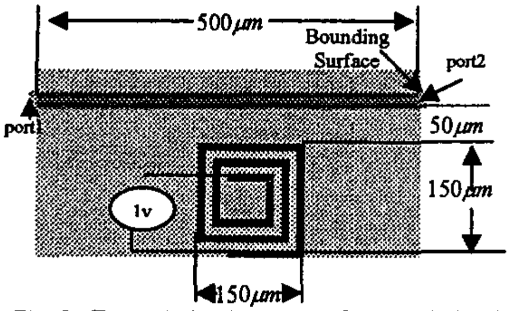


Fig. 2a Transmission line near a 3 turn spiral inductor, $10\mu m$ above a ground plane. The inductor turns are $10\mu m$ wide with $10\mu m$ gaps between the turns. The conductors of the transmission line are $3\mu m$ wide with a $10\mu m$ gap for DESIGN1, and $5\mu m$ wide with a $8\mu m$ gap for DESIGN2

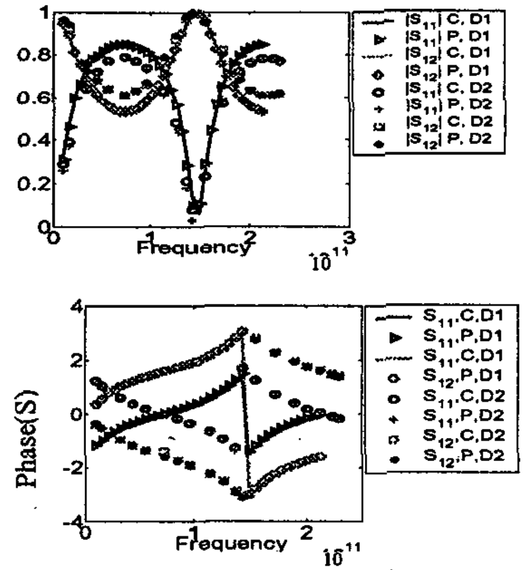


Fig. 2b Scattering parameter for DESIGN1(D1) and DESIGN2(D2), using the fully coupled method (C) and the proposed method (P).

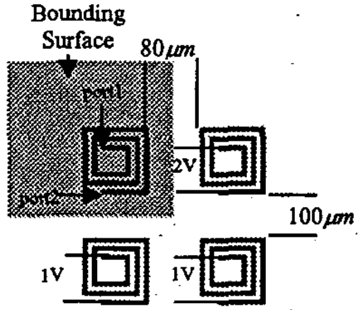


Fig. 3a Four 3 turn $150\mu m \times 150\mu m$ spiral inductors The inductor turn are $10\mu m$ wide with $10\mu m$ gaps between the turns.

Freq		S_{11}	S_{12}	S_{21}	S_{22}
60 GHz	1	$.98\angle -11^0$	$.16\angle -102^0$	$.16\angle -102^0$	$.98\angle -11^0$
	2	$.98\angle -12^0$	$.15\angle -101^0$	$.17\angle -101^0$	$.98\angle -11^0$
	3	$.98\angle -12^0$	$.15\angle -101^0$	$.17\angle -101^0$	$.98\angle -11^0$
100 GHz	1	$.99\angle -65^0$	$.08\angle -156^0$	$.08\angle -156^0$	$.99\angle -65^0$
	2	$.96\angle -69^0$	$.04\angle -128^0$	$.09\angle -168^0$	$1.0\angle -61^0$
	3	$.96\angle -70^0$	$.04\angle -127^0$	$.09\angle -169^0$	$1.0\angle -61^0$

Table1. Scattering parameters of the top-left inductor in Fig.3 using the fully coupled analysis (2), proposed isolation technique (3) and ignoring the coupling effects (1)

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